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Editor: F. A. Sowan Assistant Editor: E. G. Evans
Central Technical Services,
Mullard Limited, Mullard House, Torrington Place, London, WCIE 7HD
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Practical transient suppression circuits for thyristor power-control systems

J. MERRETT
Mullard Central Application Laboratory

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Information on the transients to be encountered on mains supplies is given as an additional aid to the design of input suppression filters. Practical suppression circuits for the commonly used thyristor control circuits with various types of load are described.

INTRODUCTION

The thyristors in a power-control system must be protected against the effects of voltage transients to prevent spurious turn-on. When the thyristors in a control system are turned on by means other than the gate signals, the unwanted flow of current may damage the load or the thyristors themselves, or constitute a hazard to personnel. The transients that can cause this spurious turn-on either have a voltage peak greater than the forward breakover voltage V_{BO} of the thyristor, or have a rate of rise greater than the dV/dt rating of the thyristor.

The nature and production of transients that can affect thyristors was considered in some detail in a previous article (Ref. 1). It was shown that these transients could be removed or reduced to a safe level by an input filter and individual device suppression circuits. A design procedure was outlined for the input filter, and design abacs derived. Some general recommendations on the degree of suppression for various types of load were made.

In this article more specific recommendations are made, and practical suppression circuits for different loads are given.

TYPES OF TRANSIENT

For convenience, the transients that can affect the operation of thyristors can be classified into three groups. These are: transients on the mains supply itself, transients produced by the operation of the mains contactor, and transients produced inside the thyristor stack by devices turning on or off. These three groups of transients were discussed in Ref. 1, and it was shown that two types of suppression circuit could be used, either alone or in combination. These circuits were a shunt-capacitance input filter and a series combination of resistor and capacitor across each thyristor. The individual device suppression circuit controls the transients produced by thyristor switching, and in combination with the input filter suppresses contactor transients. The input filter, sometimes used with a series line choke, suppresses supply voltage transients.

Supply voltage transients

The design procedure for the input filter given in Ref. 1 resulted in two design abacs, one for 600V thyristors and the other for 1200V thyristors. These two ratings are the ones most commonly used for single-phase supplies and three-phase supplies respectively. Each abac relates the maximum peak voltage of the transient to its duration for various input filters at the particular voltage rating. For convenience, the abac for 600V thyristors is shown again in Fig. 1. The value of shunt capacitance in the input filter is 4μF in all cases, and the lines A to D represent different inductance values, the sum of the inductance of the supply mains and the series line choke. Thus for a transient with a peak voltage of 500V and a duration of 10μs, it will be necessary to use an input filter with a combined mains and choke inductance of 200μH to ensure safe operation of 600V thyristors.

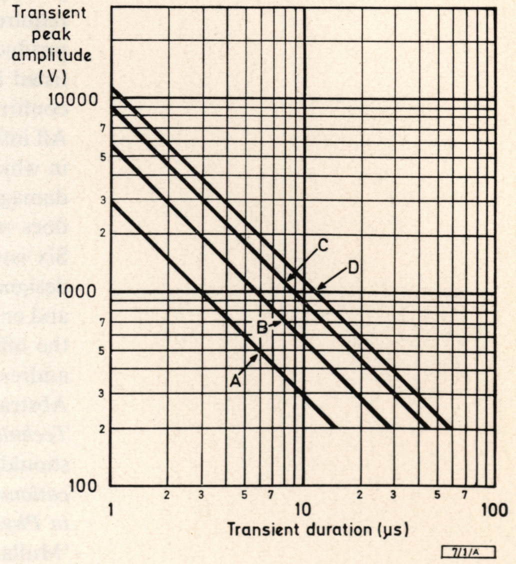

Fig. 1—Maximum permissible transient peak voltage-time product for single-phase 240V circuits using 600V thyristors

Curve	Inductance (μH)	Capacitance (μF)	Q-factor
A	50	4	∞
B	200	4	∞
C	500	4	∞
D	500	4	5

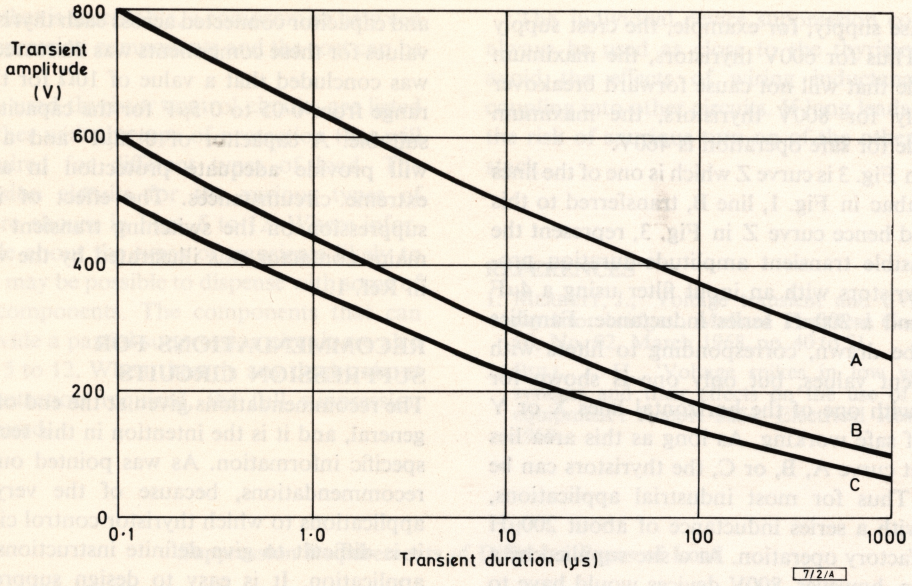

Fig. 2—Transient peak amplitude plotted against duration for three mains supplies

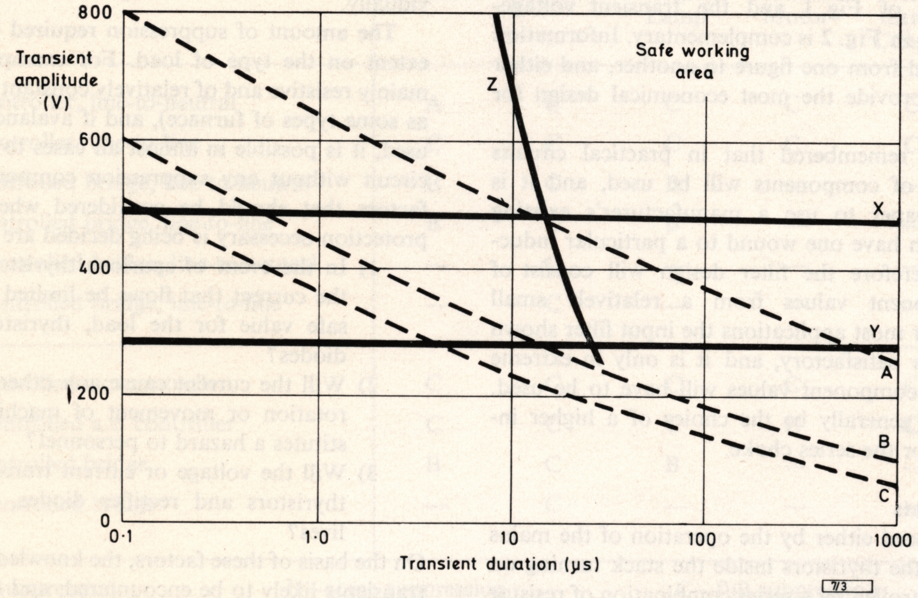

Fig. 3—Transient peak amplitude-duration curves with filter design curves superimposed

To ensure that the filter design based on this abac is the most economical one, further information on the supply voltage transients is required. This is provided by the curves in Fig. 2. Curves A, B, and C represent the peak voltage and duration of transients at three types of location. These are: a remote site served by a long low-voltage overhead line (Curve A), an industrial site served by a cable from a substation transformer (Curve B), and a site supplied by a low-voltage distribution cable (Curve C). Details of the sites and measurements are given in Ref. 2.

Strictly speaking, these curves should be bands as they represent statistical averages. The curves shown, however, may be used for design purposes.

The curve corresponding to the type of site for a particular application can be transferred to the design abac in Fig. 1. As long as the transferred curve lies below a line for a particular input filter, safe operation of thyristors with that input filter will result.

The information contained in Fig. 2 can be used directly for the design of input filters. In Fig. 3 the three curves A, B, and C are shown again with two horizontal lines superimposed. These two lines represent the maximum voltage level of transients for the safe operation of particular thyristors, line X for 800V devices and line Y for 600V devices. The voltage level for the maximum transient amplitude is the difference between the forward breakover voltage of the thyristor and the crest supply voltage. For a 240V single-phase supply, for example, the crest supply voltage is 340V. Thus for 600V thyristors, the maximum transient amplitude that will not cause forward breakover is 260V. Similarly for 800V thyristors, the maximum transient amplitude for safe operation is 460V.

Also shown on Fig. 3 is curve Z which is one of the lines from the design abac in Fig. 1, line B, transferred to this figure. Line B, and hence curve Z in Fig. 3, represent the maximum permissible transient amplitude-duration product for 600V thyristors with an input filter using a 4μF shunt capacitor and a 200μH series inductance. Families of curves could be drawn, corresponding to filters with different component values, but only one is shown for clarity. Curve Z with one of the horizontal lines X or Y defines an area of safe working. As long as this area lies above the relevant curve A, B, or C, the thyristors can be operated safely. Thus for most industrial applications, 600V thyristors with a series inductance of about 200μH will provide satisfactory operation. At a site supplied by a long overhead line, however, 800V devices would have to be used, or the values of shunt capacitance and series inductance increased.

It can be seen, therefore, that the information given in the design abac of Fig. 1 and the transient voltage-duration product in Fig. 2 is complementary. Information can be transferred from one figure to another, and either can be used to provide the most economical design for the input filter.

It should be remembered that in practical circuits preferred values of components will be used, and it is considerably cheaper to use a manufacturer's existing choke rather than have one wound to a particular inductance value. Therefore the filter design will consist of choosing component values from a relatively small range. In fact, for most applications the input filter shown in Fig. 4 will be satisfactory, and it is only in extreme cases that other component values will have to be used. The change will generally be the choice of a higher inductance value for the series choke.

Switching transients

Transients produced either by the operation of the mains contactor or by the thyristors inside the stack turning on or off can be controlled by a series combination of resistor and capacitor connected across each thyristor. The range of values for these components was discussed in Ref. 1, and it was concluded that a value of 10Ω for the resistor and a range from 0·05 to 0·5μF for the capacitor were the most suitable. A capacitor of 0·22μF and a resistor of 10Ω will provide adequate protection in all but the most extreme circumstances. The effect of individual device suppression on the switching transient produced by the mains contactor was illustrated by the waveforms shown in Ref. 1.

RECOMMENDATIONS FOR SUPPRESSION CIRCUITS

The recommendations given at the end of Ref. 1 were very general, and it is the intention in this section to give more specific information. As was pointed out in the previous recommendations, because of the very wide range of applications to which thyristor control circuits can be put, it is difficult to give definite instructions about a specific application. It is easy to design suppression circuits to give complete protection regardless of cost, but more difficult to design to an adequate level of protection at minimum cost. Each application must be assessed individually.

The amount of suppression required depends to some extent on the type of load. For example, if the load is mainly resistive and of relatively constant impedance (such as some types of furnace), and if avalanche thyristors are used, it is possible in almost all cases to use the thyristor circuit without any suppression components at all. The factors that should be considered when the degree of protection necessary is being decided are listed below.

1) In the event of spurious thyristor triggering, will the current that flows be limited by the load to a safe value for the load, thyristors, and rectifier diodes?
2) Will the current cause any other damage such as rotation or movement of machinery which constitutes a hazard to personnel?
3) Will the voltage or current transients damage the thyristors and rectifier diodes, or shorten their lives?

On the basis of these factors, the knowledge of the types of transients likely to be encountered, and the degree of risk of a device failure that is acceptable, a compromise between the number of suppression components and the cost can be reached.

The commonly used thyristor control circuits are listed in the table, together with the type of protection that will generally be required for different types of load. The transient suppression circuits for the various types of control circuits are shown in Figs. 5 to 12. Where information is available about the types of transients likely to be encountered, it may be possible to dispense with some of the suppression components. The components that can be omitted to provide a partial suppression circuit are also indicated in Figs. 5 to 12. When there is any doubt about the degree of protection required, the full suppression circuit should be used.

The individual device suppression components should always be used as close to the thyristors as possible to avoid the effects of wiring inductance and inductive coupling into other circuits. If long leads are used, there is the risk of spurious turn-on of the other thyristors in the stack.

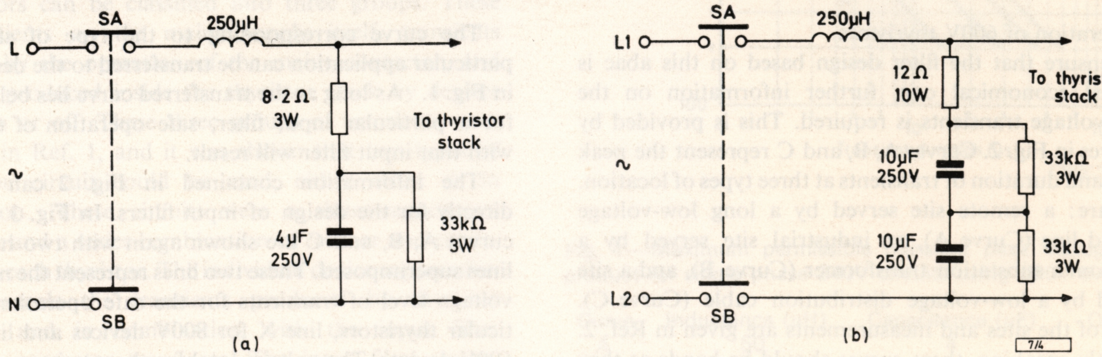

Fig. 4—Transient suppression input filter with most commonly used component values
(a) line-to-neutral (b) line-to-line

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2. BULL, J. H., 'Voltage spikes in low voltage distribution systems and their effects on the use of electronic control equipment'. Report 5254, Electrical Research Association, 1968.

Suppression Requirements of Different Types of Load

	Stack Configuration	Type of Load Resistive	Inductive	Tungsten Lamp	Induction Motor	Trans-former Coupled	D.C. Shunt Motor
Single Phase	A.C. controller, line-to-neutral	A	C	C	C	C	—
	A.C. controller, line-to-line	C	C	C	C	C	—
	Half-controlled bridge, line-to-neutral	B	C	B	—	—	B
	Half-controlled bridge, line-to-line	B	C	B	—	—	C
	Fully controlled bridge, line-to-neutral	—	C	—	—	—	C
	Fully controlled bridge, line-to-line	—	C	—	—	—	C
Three Phase	Half-controlled a.c. controller	C	C	C	C	C	—
	Fully controlled a.c. controller	C	C	C	C	C	—
	Half-controlled bridge	B	C	B	—	—	B
	Fully controlled bridge	—	C	—	—	—	C

A—no suppression B—some suppression C—full suppression

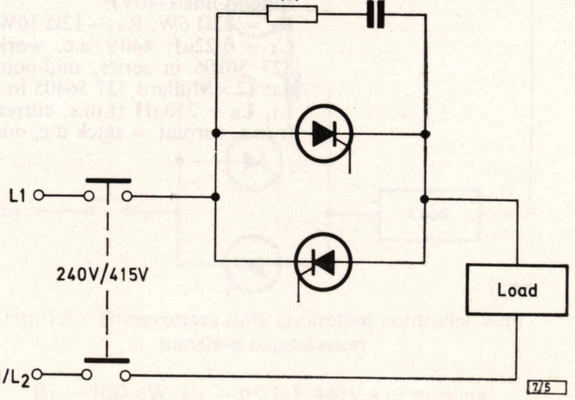

Fig. 5—Single-phase a.c. controller with transient suppression

Line-to-neutral (240V):
$R_1 = 10\Omega$ 2W, $C_1 = 0{\cdot}22\mu F$ 250 V a.c. working.

Line-to-line (440V):
$R_1 = 10\Omega$ 6W, $C_1 = 0{\cdot}22\mu F$ 440V a.c. working.

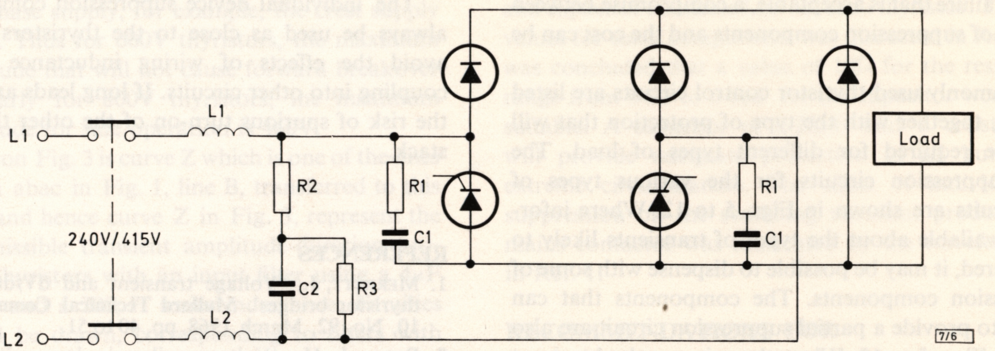

Fig. 6—Single-phase half-controlled bridge with full transient suppression for resistive and motor loads

Line-to-neutral (240V):
$R_1 = 10\Omega$ 2W, $R_2 = 8{\cdot}2\Omega$ 3W, $R_3 = 33k\Omega$ 3W;
$C_1 = 0{\cdot}22\mu F$ 250V a.c. working, $C_2 = 4\mu F$ 250V a.c. (Mullard 327 56405);
$L_1 = 250\mu H$ (r.m.s. current = stack d.c. output current), L_2 not required.

Line-to-line (440V):
$R_1 = 10\Omega$ 6W, $R_2 = 12\Omega$ 10W, $R_3 = 2 \times 33k\Omega$ 3W in series;
$C_1 = 0{\cdot}22\mu F$ 440V a.c. working, $C_2 = 2 \times 10\mu F$ 250V a.c. (2 × Mullard 327 56106 in series, mid-point to mid-point of R_3);
$L_1, L_2 = 250\mu H$ (r.m.s. current = stack d.c. output current).

Partial suppression for both cases: omit components R_1 and C_1.

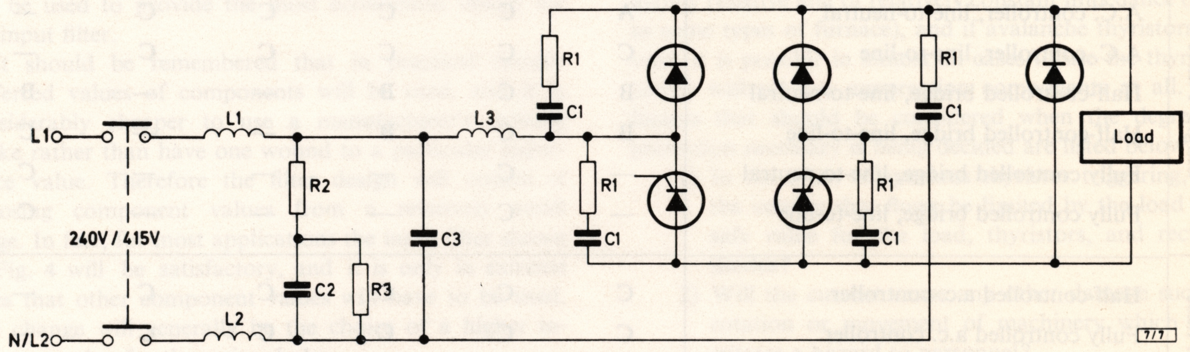

Fig.. 7—Single-phase half-controlled bridge with full transient suppression for highly inductive loads

Line-to-neutral (240V):
$R_1 = 22\Omega$ 2W, $R_2 = 8{\cdot}2\Omega$ 3W, $R_3 = 33k\Omega$ 3W;
$C_1 = 0{\cdot}22\mu F$ 250V a.c. working, $C_2 = 4\mu F$ 250V a.c. (Mullard 327 56405),
$C_3 = 1{\cdot}5\mu F$ 250V a.c. (Mullard 327 56155);
$L_1 = 250\mu H$ (r.m.s. current = stack d.c. output current), L_2 not required,
$L_3 = 50\mu H$ (r.m.s. current = stack d.c. output current).

Line-to-line (440V):
$R_1 = 22\Omega$ 6W, $R_2 = 12\Omega$ 10W, $R_3 = 2 \times 33k\Omega$ 3W in series;
$C_1 = 0{\cdot}22\mu F$ 440V a.c. working, $C_2 = 2 \times 10\mu F$ 250V a.c. (2 × Mullard 327 56106 in series, mid-point to mid-point of R_3), $C_3 = 2 \times 4\mu F$ 250V a.c. (2 × Mullard 327 56405 in series);
$L_1, L_2 = 250\mu H$ (r.m.s. current = stack d.c. output current), $L_3 = 100\mu H$ (r.m.s. current = stack d.c. output current).

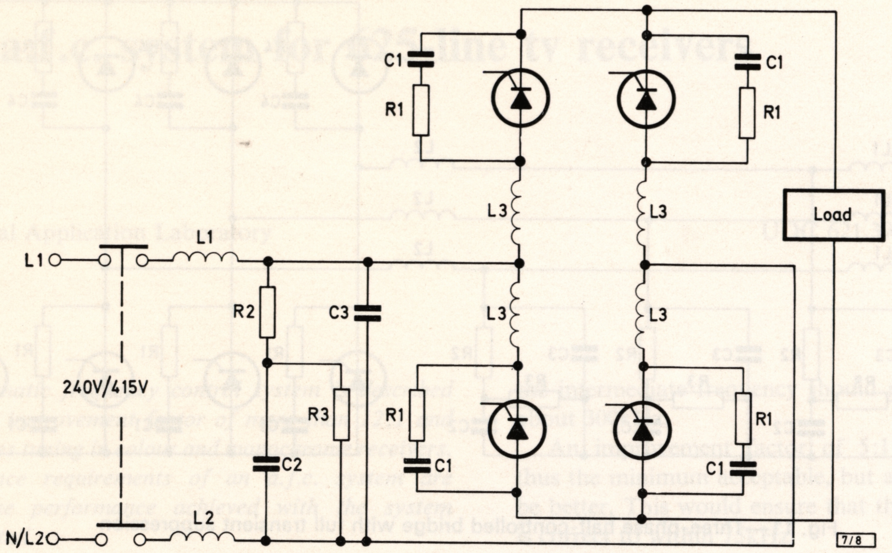

Fig. 8—Single-phase fully controlled bridge with full transient suppression

Line-to-neutral (240V):
$R_1 = 47\Omega$ 2W, $R_2 = 8{\cdot}2\Omega$ 3W, $R_3 = 33k\Omega$ 3W;
$C_1 = 0{\cdot}1\mu F$ 250V a.c. working, $C_2 = 4\mu F$ 250V a.c. (Mullard 327 56405),
$C_3 = 1{\cdot}5\mu F$ 250V a.c. (Mullard 327 56155);
$L_1 = 250\mu H$ (r.m.s. current = stack d.c. output current), L_2 not required,
$L_3 = 250\mu H$ (r.m.s. current = $0{\cdot}7 \times$ stack d.c. output current).

Line-to-line (440V):
$R_1 = 68\Omega$ 6W, $R_2 = 12\Omega$ 10W, $R_3 = 2 \times 33k\Omega$ 3W in series;
$C_1 = 0{\cdot}1\mu F$ 440V a.c. working, $C_2 = 2 \times 10\mu F$ 250V a.c. (2 × Mullard 327 56106 in series, mid-point to mid-point of R_3), $C_3 = 2 \times 4\mu F$ 250V a.c. (2 × Mullard 327 56405 in series);
$L_1, L_2 = 250\mu H$ (r.m.s. current = stack d.c. output current), $L_3 = 500\mu H$ (r.m.s. current = $0{\cdot}7 \times$ stack d.c. output current).

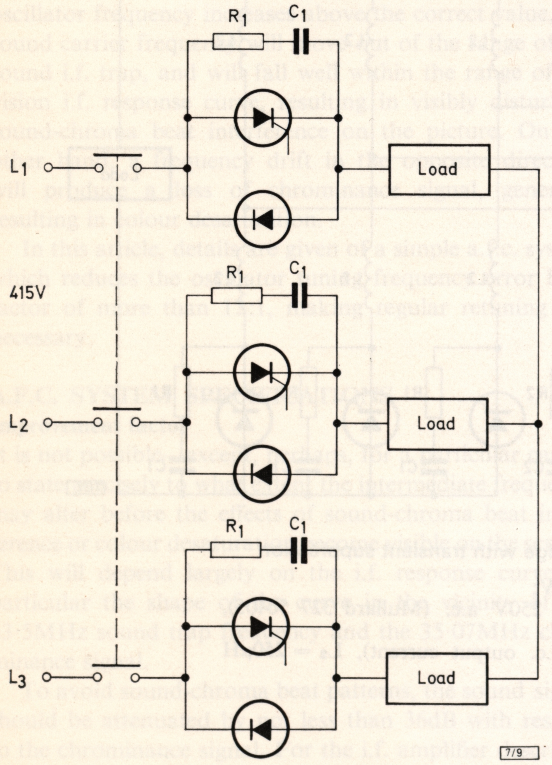

Fig. 9—Three-phase half-controlled controller with transient suppression

$R_1 = 10\Omega$ 6W, $C_1 = 0{\cdot}22\mu F$ 440V a.c. working.

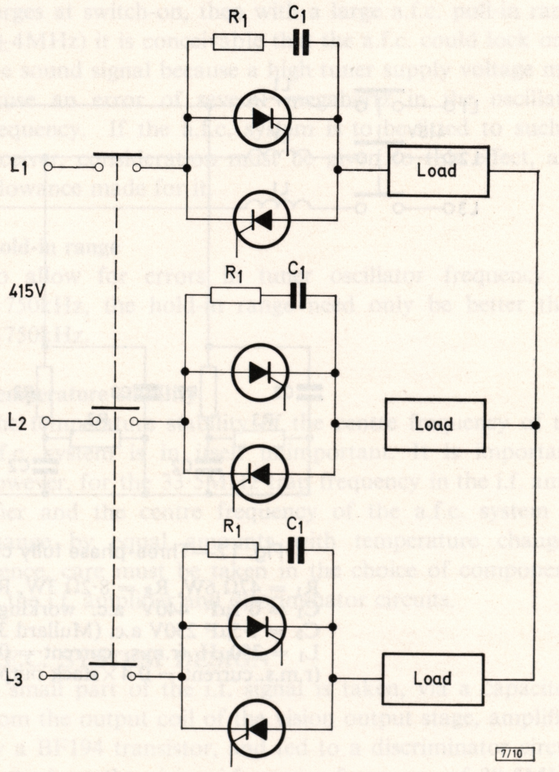

Fig. 10—Three-phase fully controlled controller with transient suppression

$R_1 = 10\Omega$ 6W, $C_1 = 0{\cdot}22\mu F$ 440V a.c. working.

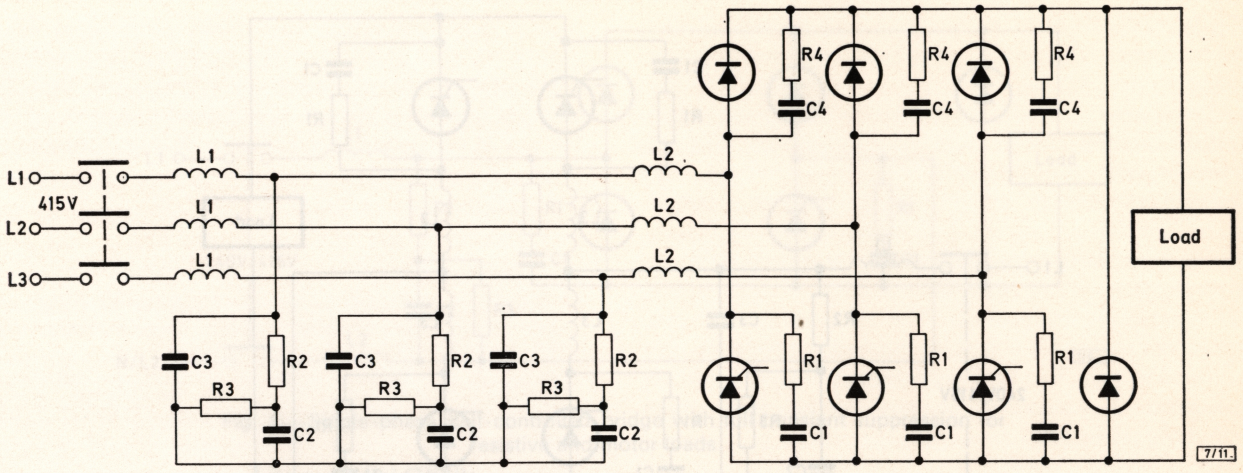

Fig. 11—Three-phase half-controlled bridge with full transient suppression

R_1, R_4 = 22Ω 6W, R_2 = 8·2Ω 3W, R_3 = 33kΩ 3W;
C_1, C_4 = 0·22μF 440V a.c. working, C_2 = 4μF 250V a.c. (Mullard 327 56405),
C_3 = 1·5μF 250V a.c. (Mullard 327 56155);
L_1 = 250μH (r.m.s. current = 0·58 × stack d.c. output current), L_2 = 50μH (r.m.s. current = 0·58 × stack d.c. output current).
Partial suppression: omit components R_4, C_3 and C_4, L_2.

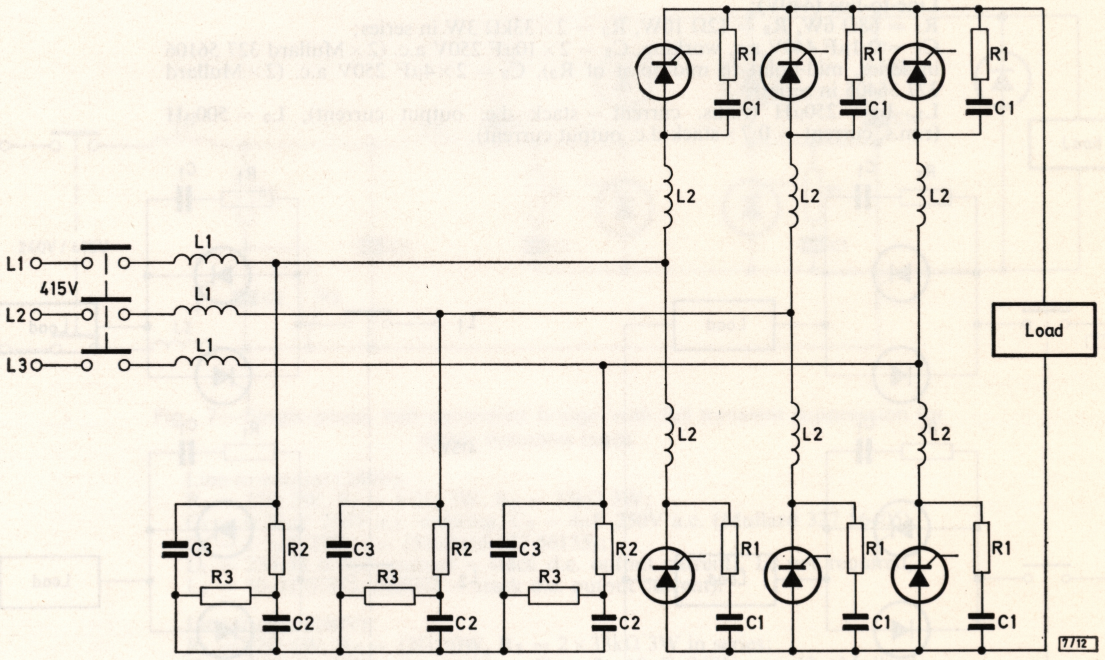

Fig. 12—Three-phase fully controlled bridge with transient suppression

R_1 = 47Ω 6W, R_2 = 8·2Ω 3W, R_3 = 33kΩ 3W;
C_1 = 0·1μF 440V a.c. working, C_2 = 4μF 250V a.c. (Mullard 327 56405),
C_3 = 1·5μF 250V a.c. (Mullard 327 56155);
L_1 = 250μH (r.m.s. current = 0·58 × stack d.c. output current), L_2 = 250μH (r.m.s. current = 0·4 × stack output current).

Simple a.f.c. system for 625-line tv receivers

P. BISSMIRE
Mullard Central Application Laboratory

UDC 621.397.132:621.316.726.078

A simple automatic frequency control system is described which gives an improvement factor of more than 12:1 and greatly simplifies tuning in colour and monochrome receivers. The performance requirements of an a.f.c. system are stated and the performance achieved with the system described in detail.

INTRODUCTION

The local oscillator in a television tuner is subject to frequency drift resulting from changes in supply voltage and temperature. In addition, when a mechanical push-button unit is used for channel selection, there may be considerable reset error. During u.h.f. operation, it is possible to have a frequency drift of some ±750kHz from a combination of the above effects.

A change of this magnitude in the resulting intermediate frequency may not be intolerable for 625-line operation in a monochrome receiver, but in a colour receiver periodic readjustment of the fine tuning would be essential. If the oscillator frequency increases above the correct value, the sound carrier frequency will move out of the range of the sound i.f. trap, and will fall well within the range of the vision i.f. response curve, resulting in visibly disturbing sound-chroma beat interference on the picture. On the other hand, a frequency drift in the opposite direction will produce a loss of chrominance signal, generally resulting in colour desaturation.

In this article, details are given of a simple a.f.c. system which reduces the oscillator tuning frequency error by a factor of more than 12:1, making regular retuning unnecessary.

A.F.C. SYSTEM SPECIFICATIONS

Improvement factor

It is not possible—except, perhaps, for a particular case—to state precisely to what extent the intermediate frequency may alter before the effects of sound-chroma beat interference or colour desaturation become visible on the screen. This will depend largely on the i.f. response curve, in particular the shape of the curve in the vicinity of the 33·5MHz sound trap frequency and the 35·07MHz chrominance signal.

To avoid sound-chroma beat patterns, the sound signal should be attenuated by not less than 36dB with respect to the chrominance signal. For the i.f. amplifier described in Ref. 1, and also the dual-standard circuit (Ref. 2), this will be the case if the increase in the intermediate frequency is less than 150kHz. To prevent desaturation, the intermediate frequency should not fall by more than about 300kHz.

An improvement factor of 5:1 (750kHz:150kHz) is thus the minimum acceptable, but a factor of 10:1 would be better. This would ensure that the oscillator frequency is correct to within 75kHz.

Pull-in range

With transistor tuners, the oscillator frequency at the moment of switching on is not likely to be much different from its final value. (This was not so with valve tuners.) Hence, there is no chance of the a.f.c. locking onto the wrong signal if the pull-in range is large, say, ±4MHz. The maximum likely error in tuner oscillator frequency from all factors after having been set correctly initially has been stated to be ±750kHz. Consequently, little attention need be paid to the pull-in range except to ensure that it is greater than about ±1MHz.

If, however, the receiver suffers from large voltage surges at switch-on, then with a large a.f.c. pull-in range (±4MHz) it is conceivable that the a.f.c. could lock onto the sound signal because a high tuner supply voltage may cause an error of several megahertz in the oscillator frequency. If the a.f.c. system is to be fitted to such a receiver, consideration must be given to this effect, and allowance made for it.

Hold-in range

To allow for errors in tuner oscillator frequency of ±750kHz, the hold-in range need only be better than ±750kHz.

Temperature stability

The temperature stability of the centre frequency of the a.f.c. system is in itself unimportant. It is important, however, for the 33·5MHz trap frequency in the i.f. amplifier and the centre frequency of the a.f.c. system to change by equal amounts with temperature changes. Hence, care must be taken in the choice of components in the i.f. amplifier and discriminator circuits.

A.F.C. SYSTEM DESIGN

A small part of the i.f. signal is taken, via a capacitor, from the output coil of the vision output stage, amplified by a BF194 transistor, and fed to a discriminator circuit centred on the vision i.f. carrier frequency of 39·5MHz. The resulting d.c. error signal is applied to a variable capacitance diode, type BB105, coupled into the tuner

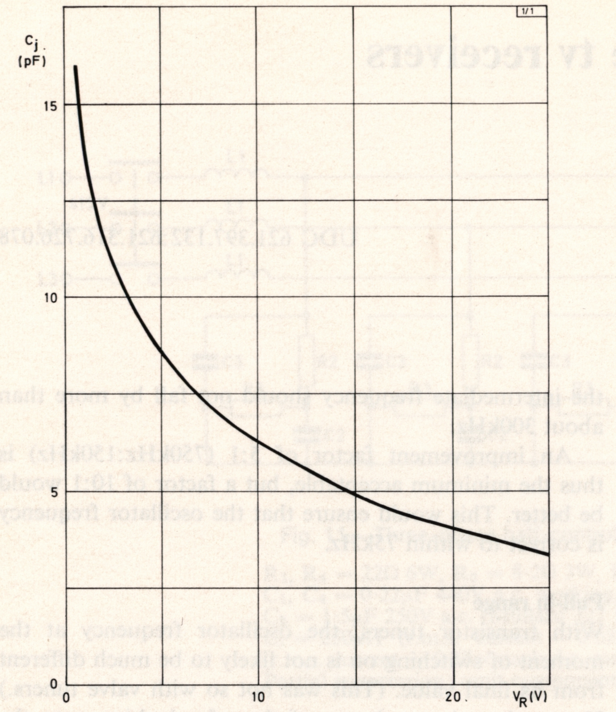

Fig. 1—Variation of junction capacitance with reverse voltage for diode BB105

oscillator circuit. The capacitance-voltage characteristic of the BB105 is shown in Fig. 1.

The tuner used in the design of the a.f.c. system is a Mullard λ/4 tuner, type AT6380/02, with the BB105 added.

Tuner oscillator control

To provide a reasonably linear relationship between applied control voltage and oscillator frequency over the entire u.h.f. tuning range (470 to 860MHz), a combination of capacitive control at the higher frequencies and inductive control at the lower frequencies is used. The method of coupling the variable capacitance diode to the oscillator circuit is shown in Fig. 2.

The diode in series with the inductance L_{126} forms a series-tuned circuit whose resonant frequency is below 470MHz. In the lower regions of the u.h.f. band, the relatively rapid change of inductive reactance of the series-tuned circuit, inductively coupled to the oscillator line by the position of L_{126}, is used to control the oscillator frequency.

At the high-frequency end of the band, the capacitance of the oscillator section of the ganged tuning capacitor is small and the tuning is effected mainly by coupling through the capacitor C_{149} tapped close to the ganged capacitor. The losses in the BB105 diode are small and consequently the a.f.c. circuit does not absorb very much power from the oscillator. Thus, provided that unduly close coupling is not attempted, the addition of such a control circuit will not necessitate redesign of the oscillator.

The two ends of the variable capacitance diode are brought out via lead-through capacitors on the tuner chassis. A nominal bias of +6·2V is applied to the diode cathode, and is derived from a potential divider fed from the tuner supply line. This value of bias enables a satisfactory frequency control range to be obtained above and below the normal centre frequency. If variations in line voltage are taken into account, the total drift with this arrangement is within 100kHz for a 10% change in supply voltage, with an improvement factor of 10:1, which is still satisfactory.

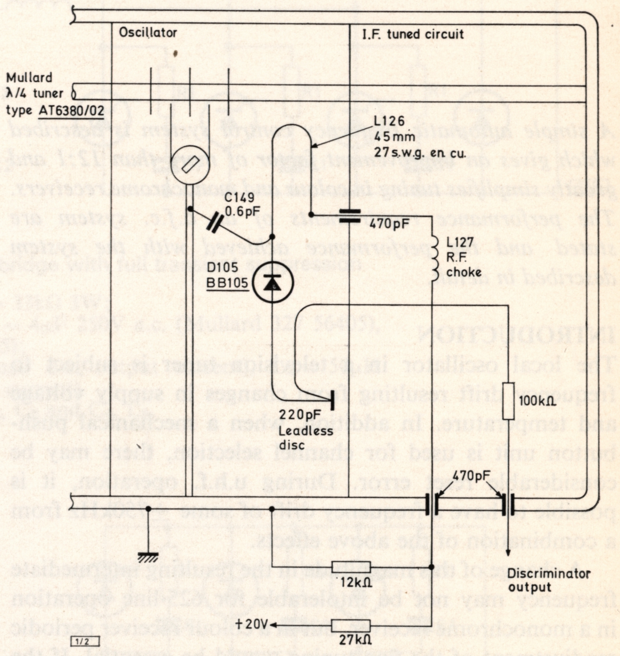
Oscillator
I.F. tuned circuit
Mullard
λ/4 tuner
type AT6380/02
L126
45mm
27 s.w.g. en. cu.
C149
0.6pF
470pF
L127
R.F.
choke
D105
BB105
220pF
Leadless
disc
100kΩ
470pF
12kΩ
+20V
27kΩ
Discriminator
output
1/2

Fig. 2—Oscillator control circuit

Discriminator circuit and amplifier

The discriminator circuit and amplifier, and also the i.f. amplifier, are shown in Fig. 3. The discriminator circuit is of basic design in which the coil Q-factors have been made as high as possible. This has been done for two reasons: firstly, a higher discriminator sensitivity (V/MHz deviation) is obtained; secondly, the pull-in range of the a.f.c. system is reduced because the discriminator skirt response is brought in closer to the centre frequency. This is shown in Fig. 4.

The design values of the discriminator circuit have been based on information from the *Radio Designer's Handbook* (Ref. 3). The primary and secondary coil Q-factors have been made equal to 40, the product kQ equal to 1·5, and the ratio of total secondary inductance to primary inductance equal to 1·77. The peak-to-peak frequency separation resulting from coil Q-factors of 40 is 1MHz.

The optimum operating conditions for the BF194 transistor have been determined experimentally by noting the peak discriminator output for a constant input at the base of the BF194. Optimum results are obtained for a collector-emitter voltage V_{CE} of 11V, and a collector current I_C of 7·3mA, which correspond to a power dissipation of 80mW. This is less than the permitted maximum for operation up to 55°C, including a derating factor of one-third.

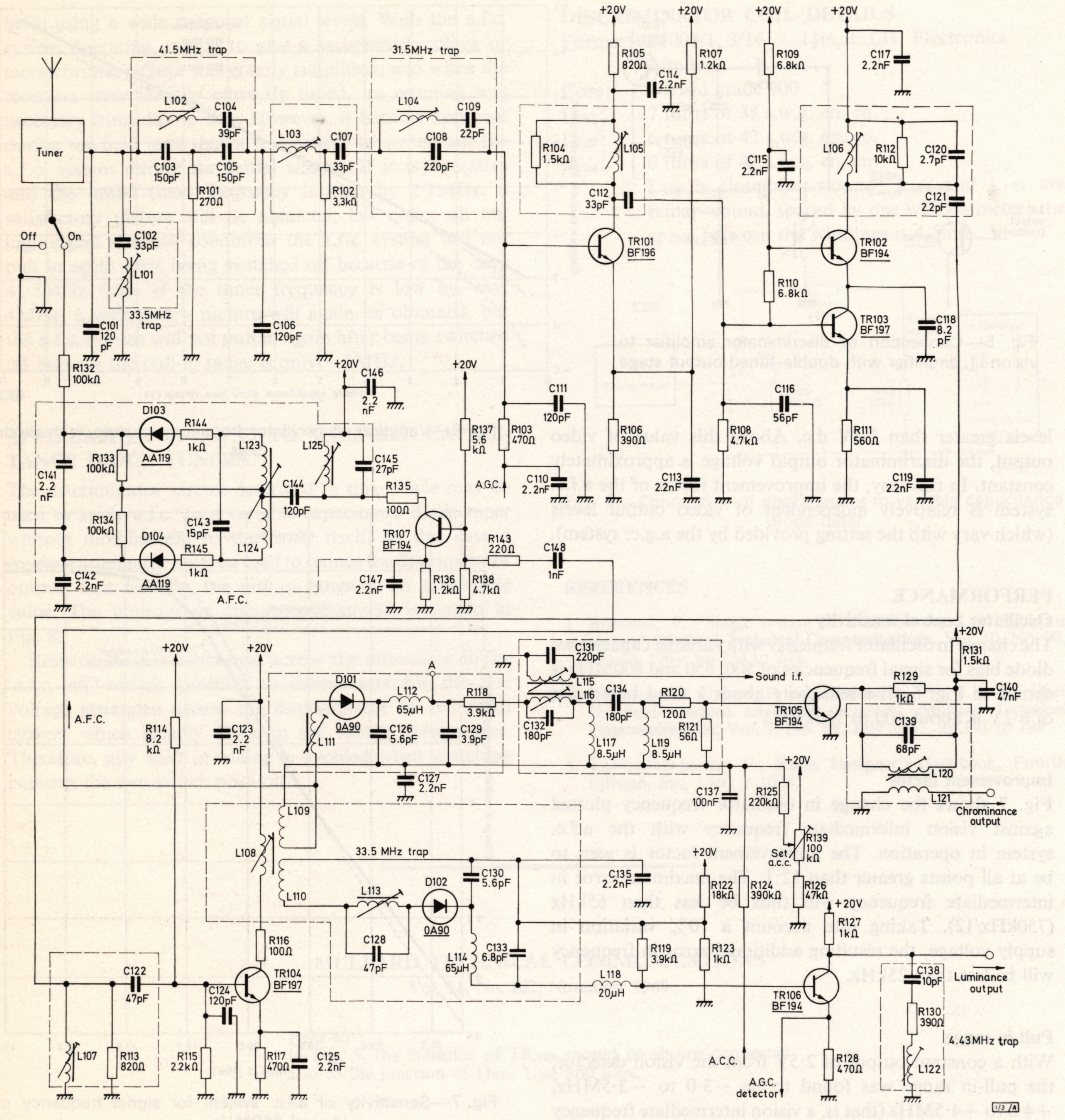
Tuner
Off
On
41.5MHz trap
31.5MHz trap
33.5MHz trap
L101
L102
L103
L104
L105
L106
C101 120 pF
C102 33pF
C103 150pF
C104 39pF
C105 150pF
C106 120pF
C107 33pF
C108 220pF
C109 22pF
R101 270Ω
R102 3.3kΩ
R104 1.5kΩ
R132 100kΩ
+20V
R105 820Ω
R107 1.2kΩ
R109 6.8kΩ
C114 2.2nF
C117 2.2nF
C112 33pF
C115 2.2nF
R112 10kΩ
C120 2.7pF
C121 2.2pF
TR101 BF196
TR102 BF194
R110 6.8kΩ
TR103 BF197
C118 8.2 pF
D103 AA119
D104 AA119
R144 1kΩ
R145 1kΩ
R133 100kΩ
R134 100kΩ
C141 2.2 nF
C142 2.2nF
C143 15pF
L123
L124
L125
C144 120pF
C145 27pF
C146 2.2 nF
R135 100Ω
TR107 BF194
R137 5.6 kΩ
C147 2.2nF
R136 1.2kΩ
R138 4.7kΩ
R143 220Ω
C148 1nF
A.F.C.
A.G.C.
C111 120pF
R103 470Ω
C110 2.2nF
R106 390Ω
C113 2.2nF
R108 4.7kΩ
C116 56pF
R111 560Ω
C119 2.2nF
A.F.C.
R114 8.2 kΩ
C123 2.2 nF
A
D101 0A90
L111
L112 65μH
C126 5.6pF
C127 2.2nF
R118 3.9kΩ
C129 3.9pF
C131 220pF
C132 180pF
L115
L116
Sound i.f.
C134 180pF
R120 120Ω
R121 56Ω
L117 8.5μH
L119 8.5μH
R131 1.5kΩ
R129 1kΩ
C140 47nF
TR105 BF194
C139 68pF
L120
L121
Chrominance output
C137 100nF
R125 220kΩ
R139 100 kΩ
Set a.c.c.
L109
L108
L110
33.5 MHz trap
L113
D102 0A90
C128 47pF
C130 5.6pF
C133 6.8pF
L114 65μH
C135 2.2nF
R122 18kΩ
R124 390kΩ
R126 47kΩ
R127 1kΩ
R119 3.9kΩ
R123 1kΩ
L118 20μH
R116 100Ω
TR104 BF197
C122 47pF
C124 120pF
L107
R113 820Ω
R115 2.2kΩ
R117 470Ω
C125 2.2nF
TR106 BF194
C138 10pF
Luminance output
R130 390Ω
4.43MHz trap
L122
R128 470Ω
A.C.C.
A.G.C. detector
1/3 /A

Fig. 3—Discriminator circuit and amplifier, and i.f. amplifier described in Ref.1

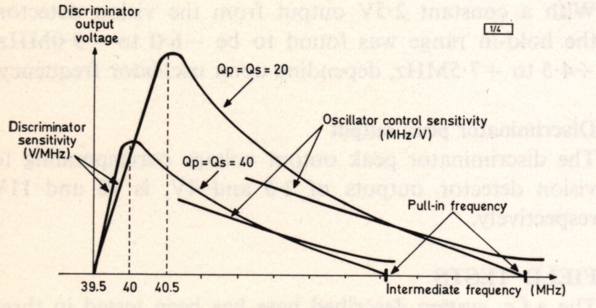

Fig. 4—Effect of coil Q-factors on pull-in frequency

The amplifier is coupled to the output stage of the vision i.f. amplifier. For the single-tuned output of the amplifier described in Ref. 1, a low-impedance, low-level feed point is available, and series damping is used to maintain stability, as shown in Fig. 3. Where a low-impedance low-level feed point is not available in the i.f. circuit, the amplifier should be coupled to the collector of the i.f. output transistor through a small capacitor, and shunt damping used to maintain stability. Fig. 5 shows this type of coupling for a double-tuned circuit.

With the coupling circuits shown, the discriminator amplifier is driven well into bottoming for video output

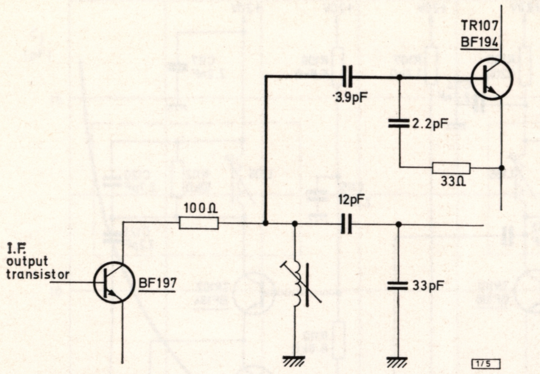

Fig. 5—Connection of discriminator amplifier to vision i.f. amplifier with double-tuned output stage

levels greater than 2·5V d.c. Above this value of video output, the discriminator output voltage is approximately constant. In this way, the improvement factor of the a.f.c. system is relatively independent of video output levels (which vary with the setting provided by the a.g.c. system).

PERFORMANCE

Oscillator control sensitivity

The change in oscillator frequency with variable capacitance diode bias for signal frequencies of 500, 650 and 800MHz is shown in Fig. 6. The sensitivity about a fixed bias point of 6·2V is about 400 to 700kHz/V.

Improvement factor

Fig. 7 shows the change in oscillator frequency plotted against vision intermediate frequency with the a.f.c. system in operation. The improvement factor is seen to be at all points greater than 12:1. The maximum error in intermediate frequency will thus be less than 65kHz (750kHz/12). Taking into account a 10% variation in supply voltage, the resulting additional error in frequency will be less than 25kHz.

Pull-in range

With a constant output of 2·5V from the vision detector, the pull-in range was found to be −3·0 to −3·5MHz, +4·0 to +4·5MHz (that is, a vision intermediate frequency of from 36 to 44MHz), depending upon oscillator frequency.

The a.f.c. system will lock-in under extreme conditions. For example, under low signal level conditions, the i.f. amplifier operates at maximum gain, giving a video output of 2·5V with the vision intermediate carrier frequency correct at 39·5MHz. If the tuner channel is changed to give a tuning error such that the vision carrier is now at (39·5 + 0·75)MHz, representing the upper limit of drift, the carrier will be attenuated by approximately 15dB, by the i.f. amplifier response relative to 39·5MHz. Thus the video output will be less than 0·5V, and the discriminator output will be similarly diminished. Measurements indicate that the discriminator output under these conditions is about 1V, and the a.f.c. will easily lock-in.

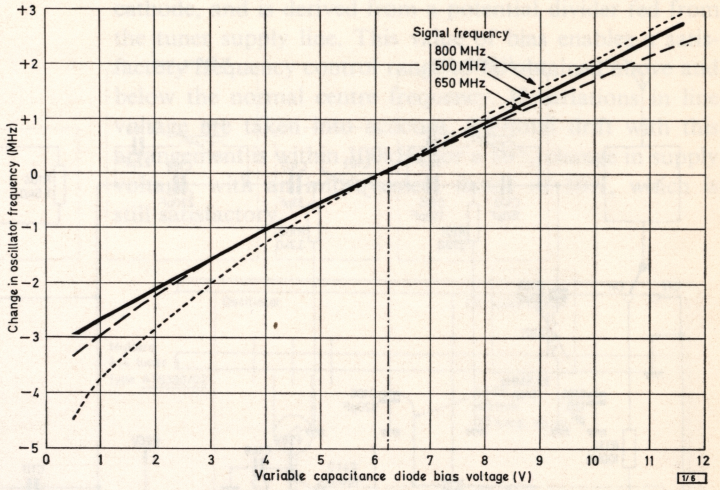

Fig. 6—Variation of oscillator frequency change with diode bias

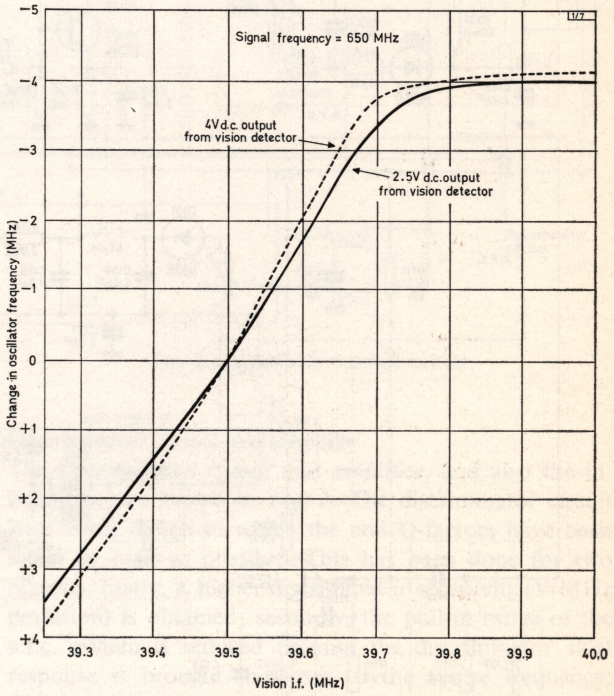

Fig. 7—Sensitivity of a.f.c. system for signal frequency of 650MHz

Hold-in range

With a constant 2·5V output from the vision detector, the hold-in range was found to be −6·0 to −9·0MHz, +4·5 to +7·5MHz, depending upon oscillator frequency.

Discriminator peak output

The discriminator peak output voltage corresponding to vision detector outputs of 2·5 and 4V, is 10 and 11V respectively.

FIELD TESTS

The a.f.c. system described here has been tested in three PAL system colour receivers in the laboratory and in the field, using a wide range of signal levels. With the a.f.c. system operating, tuning to give a satisfactory colour or monochrome picture was greatly simplified; and when the receivers were initially correctly tuned, no retuning was necessary from day to day. However, it became apparent during the field tests that it is desirable to switch out the a.f.c. system during the initial tuning. If it is operative and the initial tuner frequency is high by 2·1MHz, a satisfactory picture will be obtained, but under all but high signal strength conditions the a.f.c. system will not pull in again after being switched off because of the deep 41·5MHz trap. If the tuner frequency is low by, say, 4MHz, a satisfactory picture will again be obtained, but the a.f.c. system will not pull in again after being switched off because the pull-in range is only −3MHz.

APPLICATION OF A.F.C. TO VARIABLE CAPACITANCE DIODE TUNERS

The discriminator circuit described in this article may be used to apply a.f.c. to a variable capacitance diode tuner without modification to the tuner itself. In this case a resistive attenuator must be used to reduce the discriminator output, thus limiting the pull-in range to an acceptable value. The appropriate circuit arrangement is shown in Fig. 8.

Resistor R_{140} is connected across the attenuator in the 'a.f.c. off' switch position to compensate for the d.c. voltage generated across the discriminator by the small current which usually flows in the tuning voltage line. Therefore, any shift in tuning is avoided when switching between the two switch positions.

DISCRIMINATOR COIL DETAILS

Former:	P1003/1, 3/16 × 1½in, E.I.H. Electronics Limited
Core:	Neosid grade 900
L_{125}:	7 turns of 38 s.w.g. en. cu.
L_{123}:	6 turns of 42 s.w.g. en. cu.
L_{124}:	6 turns of 42 s.w.g. en. cu.

L_{125} is close-layer-wound; L_{123} and L_{124} are bifilar-wound, spaced by one wire diameter; the space between the windings is 4·5mm.

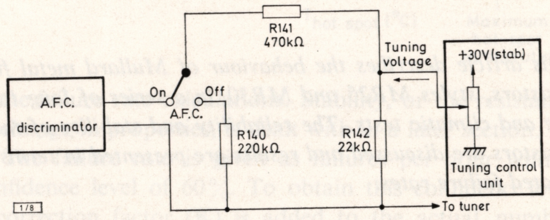

Fig. 8—Connection of discriminator to variable capacitance diode tuner

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1 Bissmire, P., 'Single system i.f. amplifier for U.K. 625-line system', **Mullard Technical Communications,** Vol. 10, No. 98, March 1969, pp.234 to 243.

2. Bridgen, R., Grant, F. and Oxenford, W.. 'I.F. amplifier for dual-standard television receivers', **Mullard Technical Communications,** Vol. 9, No. 88, July 1967, pp.191 to 199.

3. Langford-Smith, F., **Radio Designer's Handbook,** Fourth Edition, Jan. 1960, p.1020.

MULLARD TECHNICAL COMMUNICATIONS

Vol. 11, No. 102, November 1969

Page 40
In Fig. 5, the collector of TR_{203} should be shown connected also to the junction of D_{201}, C_{203} and C_{204}.

Page 41
In the seventeenth line of the left-hand column, the word 'emitter' should read 'collector'.

Reliability and stability of metal film resistors

K. W. Stanley
Industrial Electronics Division, Mullard Ltd

UDC 621.316.849:620.169

This article describes the behaviour of Mullard metal film resistors, styles MR25 and MR30, in a series of long term life and climatic tests. The reliability and stability of these resistors are discussed and results are presented in terms of quoted failure rates.

INTRODUCTION

The Mullard range of metal film resistors comprises two styles, MR25 and MR30 (Fig. 1). The method of construction of both styles is identical. The body of the resistor is a high grade ceramic rod upon which the conductor, a homogeneous film of nickel-chromium, is vacuum deposited. Solder coated, electrolytic copper connecting wires are welded to end caps of a special alloy which are pressed onto the ends of the body.

The required value of resistance is obtained by cutting a helical groove in the nickel-chromium film. The resistors are finished with at least four coats of a special lacquer and their resistance value and tolerance indicated by colour code bands.

An outline drawing and dimensions are given in Fig. 2.

The two styles of resistor, MR25 and MR30, are differentiated by their physical size, rather than by their power rating (Ref.1). The resistance ranges, limiting voltages amd maximum temperature coefficients for both types are given in Table 1.

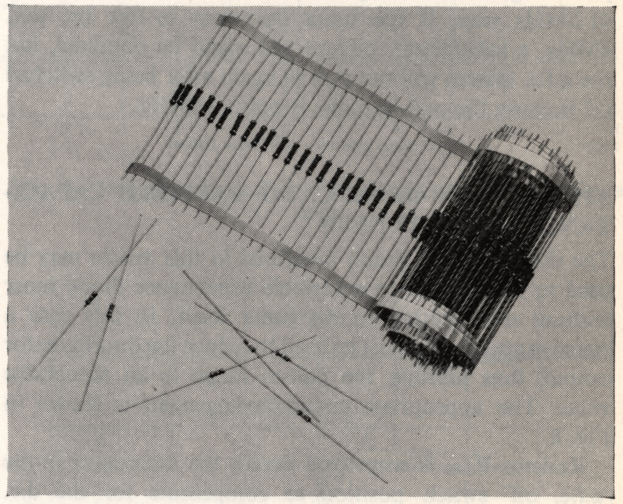
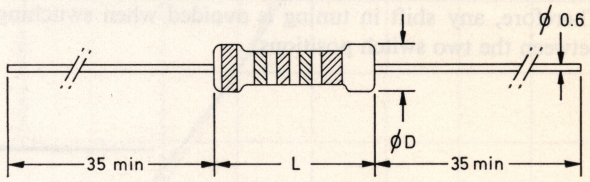


Fig. 1—Metal film resistors, styles MR25 and MR30



Style	L	D
MR25	7 max	2.5 max
MR30	10 max	3 max

All dimensions in mm

31/

Fig. 2—Outline drawing and dimensions of styles MR25 and MR30

Table 1
Type details

Style	Resistance range	Selection tolerance (%)	Series	Limiting voltage (V)	Max. temp. coefficient (×10⁻⁶ /degC)
MR25	4·99Ω to 100kΩ	1	E96	250	100
	5·1Ω to 100kΩ	2	E24	250	100
MR30	4·99Ω to 301kΩ	1	E96	350	100
	5·1Ω to 300kΩ	2	E24	350	100

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*This article is based on work done by H. Th. Dewaide and P. Popkema, Quality and Reliability Laboratory, Elcoma Division, Philips, Eindhoven.

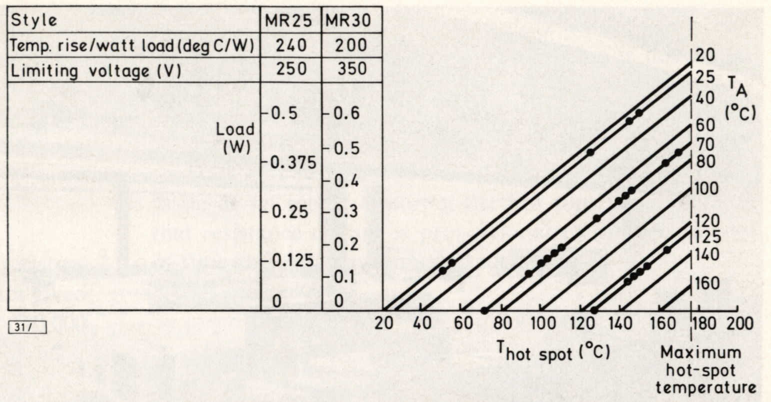


Fig. 3—Test diagram

TESTS AND MEASUREMENTS

The resistors investigated were random samples taken regularly from normal production over the period 1964 to 1968. The results can be considered representative of the entire production over that period. The main tests carried out included life, rapid change of temperature, and damp heat tests.

Life tests were performed with various combinations of temperature and loading (intermittent 1·5 hours on, 0·5 hours off), the maximum test time being 10 000 hours (see sections on failure rates and stability in long term life tests).

Rapid change of temperature tests were carried out in accordance with IEC publication number 68-2 Test Na. The conditions for these tests are 3 hours at +155°C followed by 3 hours at −55°C, this cycle being repeated five times. In these tests a failure is defined as a resistor whose value has changed by more than 0·1%.

Damp heat tests (long term exposure) were performed in accordance with IEC publication number 68-2 Test C. The resistors are loaded with 5V d.c. (except those under 10kΩ which are not loaded) and maintained at 40°C, 90 to 95% relative humidity, for 56 days. In these tests a failure is defined as a resistor whose value has changed by more than 0·5%.

RELIABILITY

The failure rate is calculated from the number of failures and the number of component test hours.

In this section only catastrophic failures are included in the failure rate calculations. Stability, or the resistance to change in properties, is dealt with in a later section.

The failure rate is given as failures per hour with a confidence level of 60%. To obtain this confidence level a correction factor (K) is added to the actual number of failures, and the failure rate (F.R.) is calculated from the expression:

$$\text{F.R. } (60\%) = \frac{\text{Number of failures} + \text{K}}{\text{Number of component hours}}.$$

The failure rate thus calculated is an average for all resistor values from the relevant production period and is only valid under the conditions in which the tests are carried out. The failure rates were further analysed, and their dependence on temperature, test time, loading, and resistor values is examined in more detail in subsequent sections.

Failure rate in long term life tests

To investigate the influence of temperature and loading on the failure rate of metal film resistors, a large number of resistors of different values were regularly tested with various combinations of temperature and loading. The temperature and loading were chosen so that samples of each type of resistor were tested at various hot spot temperatures. Fig. 3 indicates schematically which tests were performed. The combinations of temperature and loading are indicated by dots, each dot representing several types and values. The maximum test time was 10 000 hours.

In Table 2 the results are tabulated for high and low resistance values.

TABLE 2
Failure rates in life tests at various hot spot temperatures

Nominal value <100kΩ			Nominal value ⩾100kΩ		
No. tested	Component hours	No. failures	No. tested	Component hours	No. failures
9171	44 038 084	0	4743	27 907 368	12
F.R. (60%)/hour 0·02×10⁻⁶			F.R. (60%)/hour 0·49×10⁻⁶		

General view of environmental test equipment

Analysis of the results of these tests shows that failures occur only at the highest resistance values. No failures occurred at values under 100kΩ.

Table 2 shows that the failure rate for resistors of less than 100kΩ is very low, and is acceptable for higher values in spite of extreme test conditions. There was no significant difference in failure rate between the two styles.

Failure rate as a function of temperature and loading

The relationship of failure rate to temperature and loading can be determined only for resistors having nominal values greater than 100kΩ, as the tests described in the previous section produced no failures at nominal values less than 100kΩ.

It is apparent that although failure rate increases with hot spot temperature, it also depends on the combination of power dissipation and ambient temperature producing the hot spot. At identical hot spot temperatures there is a tendency for the failure rate to be greater for a high power dissipation—low ambient temperature combination than for a low power—high ambient condition. Hence, a temperature rise due to increased loading is more likely to increase the failure rate than an increase in ambient temperature. Owing to the low number of failures it is not possible to formulate the exact relationship between these factors.

Table 3 lists failures for three power dissipation—ambient temperature combinations. The greater dependence of failure rate on electrical loading is due to the increased stresses placed on thin or faulty areas of the conductive metal film which lead to excessively high localised temperatures and increased failure. Increased ambient temperatures with low loading result in equal stress over the whole resistor.

Table 3 shows that the failure rate is lower if the hot spot temperature is less than 160°C and the temperature rise due to loading does not exceed 120degC.

TABLE 3

Failure rates of metal film resistors ⩾100kΩ under various load conditions

Hot spot temperature	⩽160°C	>160°C	⩽160°C
Temp. increase due to loading	⩽120degC	⩽120degC	>120degC
No. tested	3784	767	192
Component hours	21 584 492	4 749 244	1 573 632
No. failures	4	6	2
F.R. (60%)/hour	$0{\cdot}24\times10^{-6}$	$1{\cdot}55\times10^{-6}$	$1{\cdot}97\times10^{-6}$

With hot spot temperatures above 160°C or temperature rises due to loading greater than 120degC, the failure rate is multiplied by a factor between 6 and 9.

Failure rate as a function of test time

During the entire test period the failure rate of the resistors remained practically constant. Thus, the failure rates given in the previous sections apply to the total test time of 10 000 hours.

Reliability in climatic tests

The failure data obtained from climatic tests give supplementary reliability information where these climatic conditions are relevant.

Table 4 gives a summary of the results of tests carried out.

TABLE 4

Number tested and failures in climatic tests.

	No. tested	No. failures
Rapid change in temperature	3070	0
Long term damp heat	1238	1

A total of 4308 resistors were tested and one failure was found. The estimated failure rate with a 60% confidence level is less than 0·05%.

STABILITY

In this section the stability of metal film resistors in life and climatic tests is discussed.

Stability in long term life tests

The resistors were mounted vertically and a load applied. The effect of a load applied in this way is to produce a temperature rise which is a maximum at the centre of the resistor body. The temperature rise is given by:

$$\Delta T = A.P,$$

where ΔT is the maximum temperature rise in degC, A is the temperature rise in degC/W (thermal resistance) and P is the load in watts.

The thermal resistance depends on the dimensions of the resistor, the heat conductivity of the materials used and, to a lesser degree, the manner of mounting. The sum of the maximum rise in temperature, ΔT, and the ambient temperature, T_A, is the maximum temperature, or hot spot temperature, T_M, of the resistor. Hence:

$$T_M = T_A + \Delta T = T_A + A.P.$$

The stability of a metal film resistor is determined primarily by the hot spot temperature and the resistance value. The influence of the resistance value is due to the varying thickness of metal film. Resistors having low nominal values have thicker films, and therefore greater stability, than those having high values.

Tests were made as detailed in Fig. 3, and measurements made at set intervals during the test time. It was found that resistance change is proportional to the square root of time and that the change is positive.

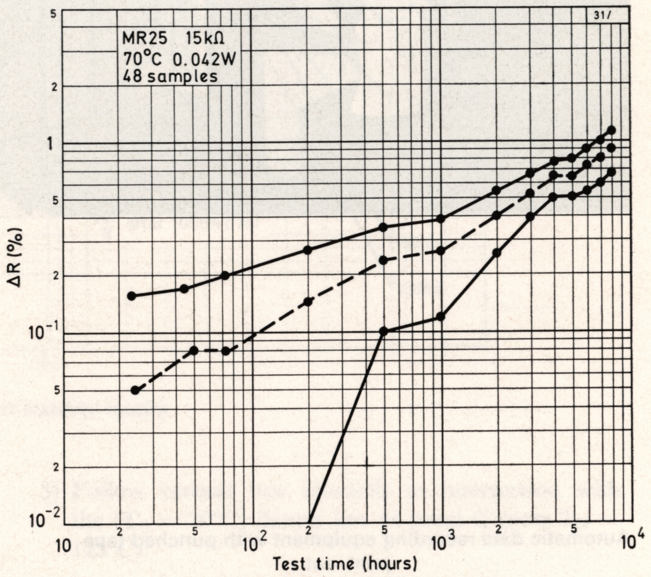


Fig. 4—Typical △R-test time characteristics

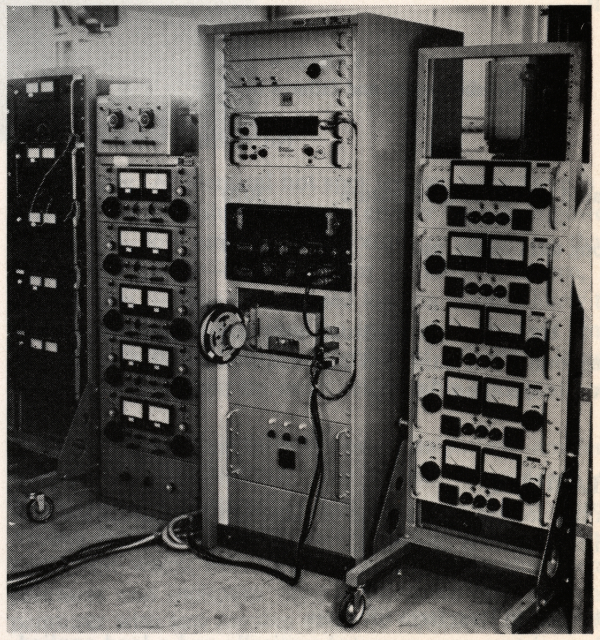
An example of the results is shown in Fig. 4. The graph shows the average resistance change and 90% interval (that is, 90% of all resistors lie between the upper and lower limits shown) plotted against time.

Measurements made by digital ohmmeter were recorded on punched paper tape and processed by computer. The principal data computed included the average resistance change, the standard deviation and the frequency distribution.

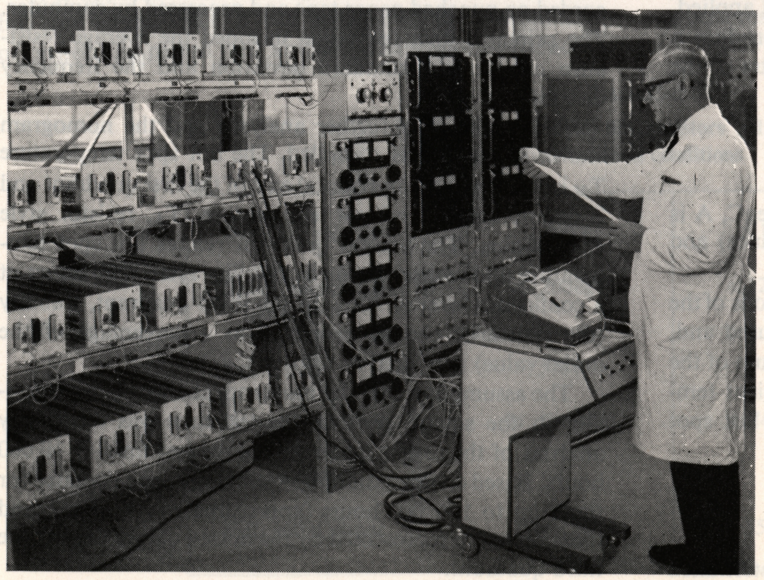
The results were further analysed and the nomogram shown in Fig. 5 was compiled. The hot spot temperature T_M is read from the horizontal axis of the first quadrant and is determined by ambient temperature, T_A, resistor style and loading. The second quadrant relates hot spot temperature and resistor value to stability after 1000 hours at a confidence level of 90%. In the third quadrant, stability (the resistance change expressed as a percentage of the initial value) is related to test time.

The assumption that the resistance changes proportionally to the square root of the test time was verified by analysis of the results taken during the test time of 10 000 hours. Based on this assumption stabilities are calculated for times up to 100 000 hours (approximately 11 years), and these times are represented by the dotted datum lines in the third quadrant.

Life testing equipment (70°C) showing portable automatic data recorder



Automatic data recording equipment with punched tape print-out



Life testing slides (room temperature) with portable automatic data recorder

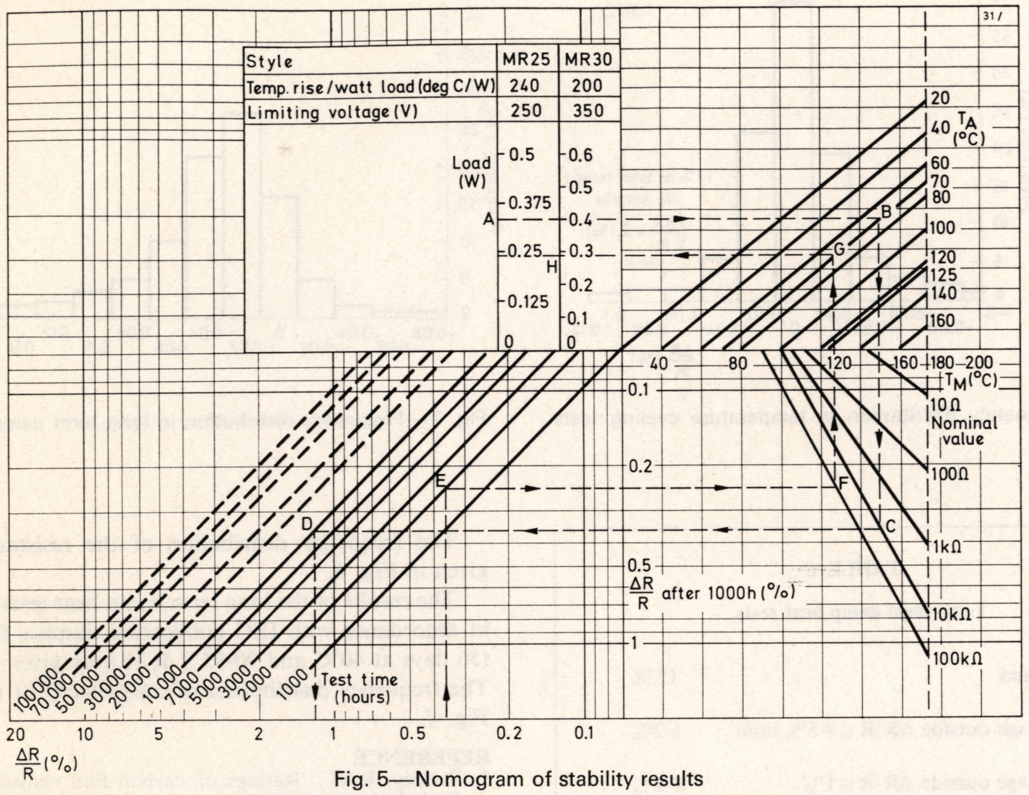


Fig. 5—Nomogram of stability results

The nomogram is used in the following way:

Example 1

Find the percentage change in resistance that occurs in a specific style and value of resistor after a number of hours' operation with defined loading and temperature; for example, style MR25, value 10kΩ, load 0·33W, ambient temperature 70°C, test time 10 000 hours.

1) Select point A (MR25, 0·33W).
2) Follow horizontal line to intersection with the (T_A = 70°C) datum line at point B.
3) Follow vertical line down to intersection with the 10kΩ datum line at point C (note T_M = 150°C).
4) Follow horizontal line to intersection with 10 000 hour test time datum line at point D (note stability after 1000 hours = 0·36%).
5) Read off stability after 10 000 hours.
   Stability = 1·15%.

Example 2

Find the maximum load permissible for a resistor of specified style and value, if the stability must be within a definite limit after a prescribed operating time and temperature; for example, style MR30, value 100kΩ, ambient temperature 70°C, operating period 2000 hours, required stability 0·35%.

1) Select point E (intersection of 0·35% stability and the 2000 hours datum line).
2) Follow horizontal line to intersection with the 100kΩ datum line at point F (note stability after 1000 hours = 0·27%).
3) Follow vertical line upwards to intersection with the (T_A = 70°C) datum line at point G (note T_M = 125°C).
4) Follow horizontal line to MR30 loading scale at point H.

The maximum permissible load which fulfils these conditions is 0·28W.

When using the nomogram the maximum value of T_M (175°C) must not be exceeded. A further limitation is imposed by the limiting voltage permissible across the resistor.

The relationship between resistance change and hot spot temperature is shown in quadrant two; the rate of resistance change with hot spot temperature varies from approximately 1·1/10degC for 10Ω to approximately 1·4/10degC for 100kΩ.

Stability in climatic tests

The results of the temperature cycling test (+155°C, 3 hours; −55°C, 3 hours; 5 cycles) are given in Table 5. Changes in resistance were required to be less than 0·1%.

TABLE 5

Temperature cycling tests — percentage outside requirements

No. tested	3070
Percentage outside $\Delta R/R \leqslant 0{\cdot}1\%$ limit.	1·1%
Percentage outside $\Delta R/R \leqslant 0{\cdot}2\%$	0·2%

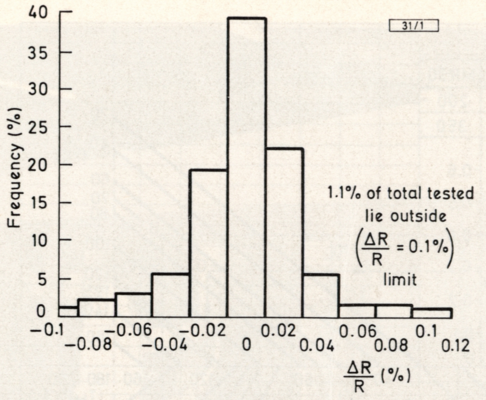


Fig. 6—Frequency distribution in temperature cycling tests

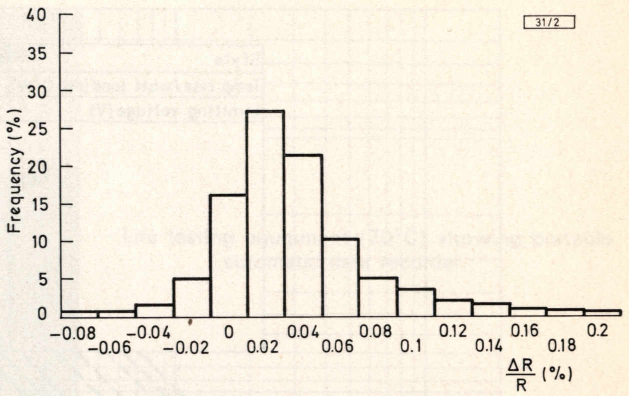


Fig. 7—Frequency distribution in long term damp heat tests

TABLE 6

Long term damp heat tests.

No. tested	1238
Percentage outside $\Delta R/R \leqslant 0{\cdot}5\%$ limit	1·2%
Percentage outside $\Delta R/R \leqslant 1\%$	0·8%

The frequency distribution of the resistance drift is given in Fig. 6.

The results of the long term damp heat tests carried out in accordance with IEC publication number 68-2 Test C (56 days at 40°C and 90-95% R.H.) are given in Table 6. The frequency distribution of resistance drift is shown in Fig. 7.

REFERENCE

1. Stanley, K.W., 'Ratings of carbon film resistors', **Mullard Technical Communications,** Vol. 9, No. 86, March 1967, pp. 144 to 148.